



# Low Drop-Out Voltage Regulator as a Candidate Topology for Photovoltaic Solar Facilities

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**Abstract.** This article aims to present the design of a 4.5-V, 450-mA low drop-out (LDO) voltage linear regulator based on a two-stage cascoded operational transconductance amplifier (OTA) as error amplifier for photovoltaic solar DC-DC regulation. The aforementioned two-stage OTA is designed with cascoded current mirroring technique to boost up the output impedance. The proposed OTA has a DC gain of 101 dB under no load condition.

The designed reference voltage included in the LDO regulator is provided by a band gap reference with the temperature coefficient ( $T_V$ ) of 0.025 mV/°C. The proposed LDO regulator has a maximum drop-out voltage of 0.5 V @ 450 mA of load current, and has the worst case power supply rejection ratio (PSRR) of [54.5 dB, 34.3 dB] @ [100 Hz, 10 kHz] in full load condition. All the proposed circuits are designed using a 0.35  $\mu$ m CMOS technology. The design is checked in order to corroborate its performance for wide range of input voltage, founding that the circuit design works fine meeting all the initial specification requirements.

## Key words

Photovoltaic solar DC-DC regulator, low drop-out (LDO) voltage linear regulator, operational transconductance amplifier (OTA), band gap reference, cascoded current mirror.

## 1. Introduction

Low-dropout voltage linear regulators (LDO) [1-4] have gained much of importance due to the increased use of battery-powered devices and photovoltaic solar DC-DC voltage regulation. The main function of the LDO regulator is to provide a reliable, stable and constant voltage, and has been considered as one of the important component in power management of cell phones, laptops, wireless applications, etc., where one of the important issues is the dropout voltage, which plays an important role in these kinds of applications [5].

Fig. 1 shows the overall topology of the proposed LDO regulator. As can be seen, it consists of three main blocks: An error amplifier, a voltage reference and a pass transistor with external load capacitance with small value of internal resistance (ESR) for the frequency compensation. The error amplifier of the LDO regulator, in the negative feedback condition, constantly compares the error signal with the reference voltage and hence maintaining the constant output voltage by varying the gate to source voltage of the pass transistor accordingly. Since the output of error amplifier is used to drive the gate of the pass transistor which is naturally a capacitive load, the best option for the error amplifier would be an operational transconductance amplifier (OTA) [6, 7] with high output impedance (Fig. 2). In addition, it is very easy to model OTA as single pole system because of its high output impedance, which forms a low frequency pole with small load capacitance. Therefore, OTAs are the best components to drive the capacitive loads because of the fact that their output signal is a current controlled by a differential input signal.

The voltage reference is one of the important blocks of the voltage regulators, since it decides the nominal output voltage. The main design issue will be its temperature coefficient, which has to be ideally zero or very near to it. There are many approaches in obtaining the voltage reference. However, one among them, with very low temperature coefficient, is the use of a band gap voltage reference (BGR) [8]. Finally, a pMOS pass transistor, which carries a major part of current in the whole circuit, drives the load. Its dimensions are set such that it is able to withstand the maximum rated current and to achieve the rated low dropout voltage.

The paper is organized as follows: In Section 2, the design of the two-stage cascoded OTA in CMOS technology is carried out. Next, in Section 3, the design of the LDO regulator is discussed, including the design of the error amplifier based on the proposed OTA, and

design of the pass transistor in CMOS technology. Finally, in Section 4, simulation results are obtained for both proposed OTA and LDO regulator.

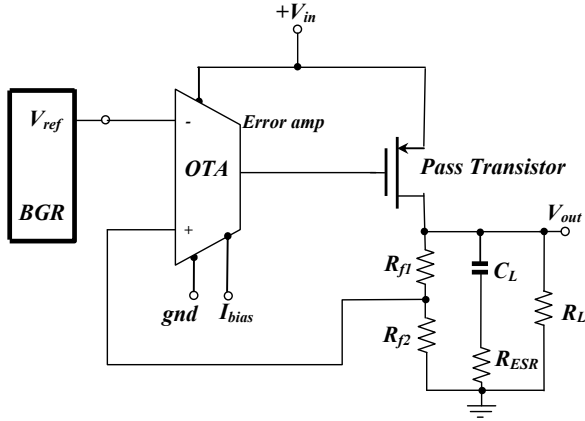


Fig. 1. Topology of the proposed LDO voltage linear regulator.

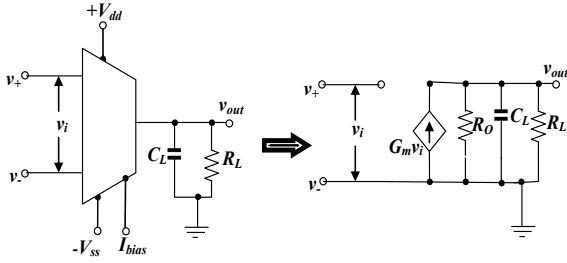


Fig. 2. Small signal model of an OTA with  $C_L$  and  $R_L$ .

## 2. Design of the Operational Transconductance Amplifier (OTA)

Fig. 3 shows the classical two-stage OTA in CMOS technology [7], where it is operated in  $+V_{dd}$  and  $-V_{ss}$  power supplies and with an external bias current  $I_{bias}$  and having a single ended output. The first stage is the normal  $n$ -channel differential input pair ( $m_1$  and  $m_2$ ) with  $p$ -channel current mirror as its active load ( $m_3$  and  $m_4$ ). The drain currents of  $m_1$  and  $m_2$  are mirrored to  $m_6$  and  $m_5$ , respectively, which is the second (gain) stage through the classical current mirroring technique with the current ratio of  $1:\alpha$ . In Fig. 3, we get:

$$I_a = (I_{bias} / 2) - g_{m(2)} v_i, \text{ and } I_b = (I_{bias} / 2) + g_{m(1)} v_i \quad (1)$$

The single ended output is taken out from point  $P1$ , through which a current of:

$$I_o = 2 \cdot g_{m(1,2)} \cdot \alpha \cdot v_i \quad (2)$$

is flowed, where  $v_i = (v_+) - (v_-)$  is called the differential input. Therefore, the voltage controlled current source is obtained and the transconductance of this OTA,  $G_m$ , is given by the expression:

$$G_m = \frac{I_o}{V_i} = 2 \cdot \alpha \cdot g_{m(1,2)} \quad (3)$$

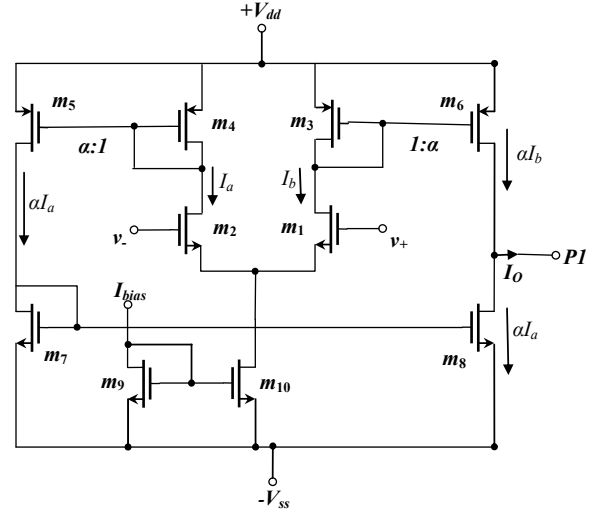


Fig. 3. Classical two-stage OTA in CMOS technology.

It should be noted that according to Equation (3), the transconductance of the OTA is dependent on the  $g_m$  of the MOS transistors of the input differential pair which is in-turn dependent on the DC current through it, i.e.,  $\sqrt{I_{bias} / 2}$ . Therefore, it can be said that it is a current controlled gain OTA. The parameter  $\alpha$  is called the gain factor which is modified by varying the ratio  $W/L$  of the second stage with respect to the input stage ( $W/L_{(6,5)} : W/L_{(3,4)}$ ). The equivalent small signal model representation of the OTA is shown in Fig. 2. According to this figure, the voltage gain ( $A_v(s)$ ) is given by the expression:

$$\frac{V_{out}(s)}{V_i(s)} = A_v(s) = \frac{A_v}{1 + \frac{s}{\omega_o}} = G_m \cdot Z_L(s) \quad (4)$$

where  $G_m$  is the transconductance of the OTA which is given by Equation (3) and the cut-off frequency  $\omega_o$  is given by:

$$\omega_o = \frac{1}{(R_o \parallel R_L) C_L} \quad (5)$$

and  $Z_L(s)$  is the output impedance of the OTA, which is given by:

$$Z_L(s) = \frac{R_o \parallel R_L}{1 + (R_o \parallel R_L) C_L(s)} \quad (6)$$

It should be noted that the OTA has the high output impedance thanks to the common source output configuration which gives the high output impedance  $r_o(m_8) \parallel r_o(m_6)$ . The main advantage of the OTA is that, due to the fact of its high output impedance, the frequency compensation is easier. The load capacitance ( $C_L$ ) will only create the dominant pole. Hence the unity gain bandwidth is varied by varying the load capacitance

to obtain the suitable phase margin (PM) according to the application and hence assuring the stability of the system.

Fig. 4 shows the improved two-stage OTA in CMOS technology, adopting the cascoded technique to boost up the output impedance and hence the DC voltage gain (no load) of the classical OTA, taking the proposed OTA into more ideality.

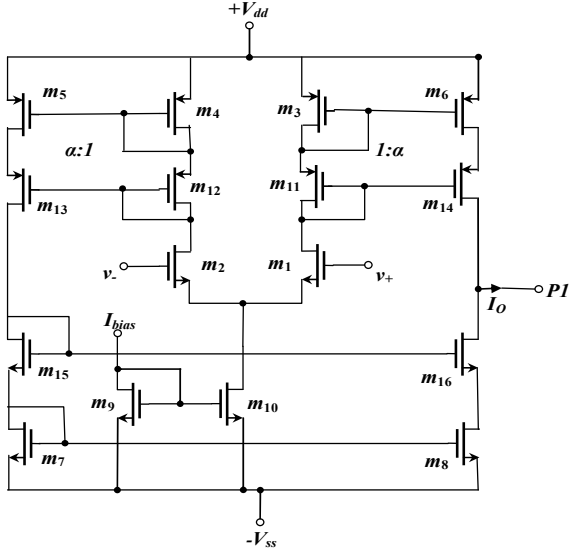


Fig. 4. Proposed two-stage cascoded OTA in CMOS technology.

In the proposed design ( $m_{4,5}$ ) and ( $m_{13,12}$ ), ( $m_{3,6}$ ) and ( $m_{11,14}$ ), ( $m_{15,16}$ ) and ( $m_{7,8}$ ) form the cascoded pairs with current mirroring. The main design criterion of the proposed OTA is same as the classical OTA design which depends on Equations (3), (4), and (6). The main advantage here is the boosting up of output impedance by a factor of approximately  $(2 + g_{m(8)} r_o(m_8))$  which is quite significant. Hence the output impedance,  $R_o$ , is varied by varying the ratios  $W/L_{(6,14,16,8)}$  in-turn varying its  $g_m$ . Regarding the other important parameters, the input impedance  $R_i \approx \infty$  (since input are at the gate of the MOSFETs), the DC voltage gain of the OTA, from Equations (4) and (6), is given by:

$$A_v = G_m (R_o \parallel R_L) \quad (7)$$

Equation (7) can be used to find the  $g_{m(1,2)}$  experimentally by obtaining the AC response of the proposed OTA for  $R_L = \infty$ , and also obtaining its output impedance plot with respect to frequency. Therefore, the equation for DC gain reduces to:

$$A_v \approx G_m R_o \quad (8)$$

In Section 4, simulation results of classical OTA are compared with proposed OTA by considering some parameters. In addition, different parameters of the proposed OTA like AC response, DC gain, output impedance, etc. are also presented.

### 3. Design of the Low Drop-Out (LDO) Voltage Linear Regulator

As mentioned in Section 1, Fig. 1 shows the classical topology of the proposed LDO regulator. It consists of three main blocks, namely *error amplifier* block, *voltage reference* block, and *pass transistor*; and it is biased by means of an external current  $I_{bias}$ . In standard voltage regulators, the main difference compared to an LDO is in the pass transistor block. Normally, in a typical voltage regulator, pass transistor element will be an *n*MOS (or *nnp*) transistor in source follower (or emitter follower) configuration which has a typical dropout voltage as  $2V$ . However, in a LDO regulator, the pass transistor element is a *p*MOS (or *pnp*) transistor open drain (or open collector) configuration, in which its dropout voltage is nothing but its saturation voltage with typical values as  $0.3 \sim 0.5V$ . One of the handicaps in a LDO regulator is, since a *p*MOS transistor is used as the pass transistor in open drain configuration (Fig. 1), it forms a high output impedance, creating a pole within the unity gain frequency. Since already there are a low frequency pole provided by error amplifier (OTA), there will be extra phase contributed by the pole formed by the high output impedance of the LDO to overall its response by decreasing the phase margin and causing the issue in stability of the system. Therefore, an output capacitance with small  $R_{ESR}$  is needed to compensate the above-mentioned pole.  $R_{ESR}$  with load capacitance ( $C_L$ ) together form a zero to compensate the effect of the pole created by the high output impedance of the LDO regulator.

Resistors  $R_{f1}$  and  $R_{f2}$  in Fig. 1 are the resistors that are used to set the output voltage ( $V_{out}$ ). The error amplifier forms a negative feedback loop, which constantly compares the error signal at the output with the reference voltage to maintain constant  $V_{out}$  by varying the gate voltage of pass transistor, and hence controlling the current flowing through it. Thus, the output voltage is given by:

$$V_{out} = \left(1 + \frac{R_{f1}}{R_{f2}}\right) V_{ref} \quad (9)$$

Thus, by choosing the appropriate values of resistors and reference voltage, output voltage  $V_{out}$  is set.

The initial design specifications given for the proposed LDO regulator are shown in Table I. Fig. 5 depicts the complete schematic of the CMOS LDO voltage linear regulator in order to fulfill the performance given by these design specifications. On the other hand, dimensions of the circuit transistors in the proposed LDO voltage linear regulator in Fig. 5 are depicted in Table II.

As can be appreciated in Fig. 5, the proposed design has the three aforementioned main parts: (1) The error amplifier, (2) the voltage reference circuit, and (3), the pass transistor and the suitable load capacitance  $C_L$ .

All the proposed circuit blocks are built in Cadence Hit-Kit V3.70, using a 0.35  $\mu\text{m}$  (c35b4c3) CMOS

technology. It is simulated using the Specter software, using the BSIM-3.3 as the MOSFET model.

#### A. Operational Transconductance Amplifier

The proposed OTA as in Fig. 4 is built and simulated with  $+V_{dd}=5\text{ V}$ ,  $-V_{ss}=-5\text{ V}$ , and with the bias current  $I_{bias} = 50\text{ }\mu\text{A}$ . The circuit is simulated in order to obtain some parameters and is compared with the results got by classical OTA shown in Fig. 3. Open loop AC Response of the proposed OTA is plotted and shown in Fig. 7 for the load capacitance  $C_L=0.4\text{ nF}$  and in the no-load condition. It should be noted that the gain is directly proportional to the

small vales of load resistance  $R_L$ . The results showed that DC gain  $A_V=101.13\text{ dB}$ ,  $UGB=4.5\text{ MHz}$  with  $PM=59.17^\circ$ .

In addition, the proposed OTA circuit is simulated with  $C_L=1\text{ nF}$ . The DC gain was same, but the  $UGB=1.98\text{ MHz}$  with  $PM=75.3^\circ$ , whereas in the classical OTA, the  $UGB=2.316\text{ MHz}$  with  $PM=81.07^\circ$ . This boosting up of the DC gain in the proposed OTA compared to classical OTA under no load conditions is due to the increase in the output impedance thanks to the cascaded current mirrors in the first one. The DC gain under no load condition matches with Equation (8).

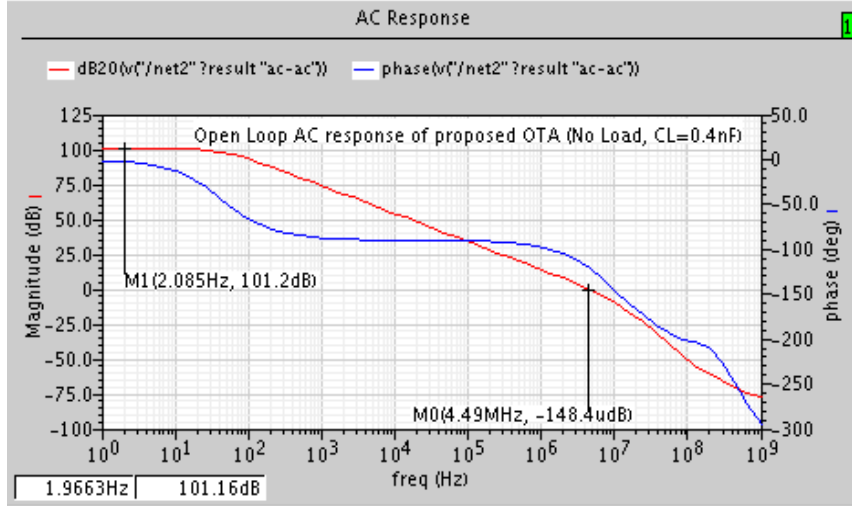


Fig. 7. AC-Response of the proposed OTA, with  $C_L=0.4\text{ nF}$ ,  $R_L=\infty$ .

Another important parameter, the output impedance, is evaluated by simulating the circuit with differential inputs grounded and applying an AC-voltage source at the output node, with  $C_L=1\text{ nF}$ , by finding the current through that voltage source. Fig. 8 shows the output impedance  $Z_L(s)$  with  $R_L=\infty$ , with respect to frequency change in proposed OTA. From this figure, it can be seen that the large signal output impedance ( $R_O$ ) in the proposed OTA is  $1.35\text{ M}\Omega$  which is a good high output impedance, thanks to the cascaded current mirroring technique in proposed OTA.

Another parameter that decides the gain is  $g_{m(1,2)}$  of the input differential pair MOSFETs. It is found out experimentally from Fig. 7 and 8, and using Equations (8) and (3) in no load condition. From these figures, the values of  $R_O$  and DC gain  $A_V$  can be obtained, with a gain factor  $\alpha$  set to 20. Substituting appropriate values, it is found that  $g_{m(1,2)}=2.12\text{ mS}$ . This value suits very well with the theoretical value found out using the model file parameters ( $k_n$ ,  $V_{tn}$ ) and DC current as  $I_{bias}/2$ , i.e.,  $2.204\text{ mS}$ .

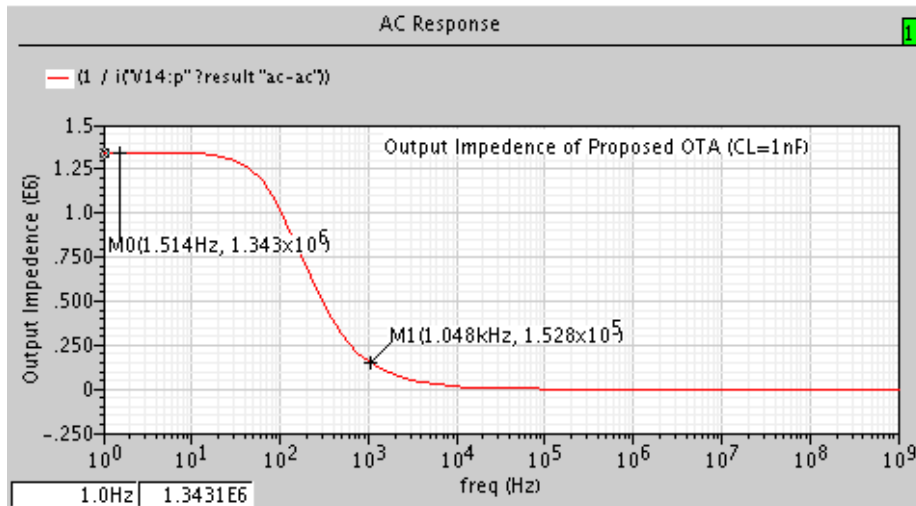


Fig. 8. Output impedance vs. frequency of the proposed OTA.

## B. LDO Voltage Regulator

Fig. 5 shows the overall proposed circuit of LDO regulator with three aforementioned main parts: The error amplifier (based on the previously proposed OTA), the band-gap reference circuit and the pass transistor. This circuit was built and simulated with  $V_{in}$  ranging from 5V to 20V,  $I_{bias}=50 \mu A$  and it is found that the designed circuit fits to the specifications mentioned in Table I. Tables III and IV show the summary of the simulation results obtained for band-gap reference voltage and LDO regulator circuits respectively.

TABLE III.- SIMULATION RESULTS OF BAND-GAP REFERENCE VOLTAGE CIRCUIT.

Parameters	Values
$V_{ref}$ at $T=25^\circ C$	653 mV
$T_\gamma$	0.025 mV/ $^\circ C$
Line Regulation	0.44 mV/V
Power drawn, $V_{in}=5 V$	0.54 mW

TABLE IV.- SIMULATION RESULTS OBTAINED FOR THE PROPOSED LDO REGULATOR.

Parameter	Value		
	For $R_L=\infty$	For $R_L=10\ \Omega$	For $R_L=45\ \Omega$
$V_{out}$ for $V_{in}=5\ V$	4.509 V	4.504 V	4.507 V
$I_L$ for $V_{in}=5\ V$	0 A	450.4 mA	100.1 mA
$V_{drop}$	0.06 V	0.5 V	0.15 V
Line regulation	3.2 mV/V	3.4 mV/V	3.2 mV/V
Load regulation	0.012 V/A		
$V_{out}$ settling time $I_L=0 - 450\ mA$ in 1 $\mu s$	4.2 $\mu s$		
$V_{out}$ settling time $I_L=450\ mA - 0$ in 1 $\mu s$	40 $\mu s$		
PSRR (full load)	@ 100 Hz	54.45 dB	
	@10 kHz	34.31 dB	
$V_{out}$ settling time $V_{in}=5\ V - 20\ V$ in 1 $\mu s$	6 $\mu s$		
$V_{out}$ settling time $V_{in}=20\ V - 5\ V$ in 1 $\mu s$	4 $\mu s$		

## 5. Conclusion

The current article has shown a proposal of low drop-out voltage regulator based on a two-stage cascaded operational transconductance amplifier (OTA) as error amplifier. It is being designed in a 0.35  $\mu m$  CMOS technology and simulated in Cadence Spectre software with the MOSFET model of BSIM -3.3.

The current measured regulated output voltage is 4.5V with a maximum current of 450 mA. The LDO regulator design has the dropout voltage of 0.5 V at full load condition, also with good line and load regulations of 3.2 mV/V and 0.012V/A, respectively. The PSRR of the design was found out to be 54.45 dB at 100 Hz and 34.31 dB at 10 kHz under full load condition. The LDO voltage linear regulator includes a band gap reference circuit that provides the constant reference voltage of 0.653 V with a temperature co-efficient ( $T_\gamma$ ) of 0.025 mV/ $^\circ C$ .

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